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IMPLICATIONS OF ADAPTIVE CANCELLATION ARRAY PROCESSING FOR DESIGN OF A SPACE-BASED SURVEILLANCE RADAR

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IMPLICATIONS OF ADAPTIVE CANCELLATION ARRAY PROCESSING FOR DESIGN OF A SPACE-BASED SURVEILLANCE RADAR

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SUMMARY

We examine the system design issues associated with the use of adaptive jammer rejection in the context of possible specifications for a space-based surveillance radar. We show that the adaptive nulling requirements of the system cannot be considered in isolation. Adaptive processing has implications for the entire system, from design of the antenna array, through to the choice of ADCs and the requirements of subsequent coherent integration and detection processing. Three different classes of adaptive processing architecture are considered: the sidelobe canceller, the fully adaptive array, and the generalised sidelobe canceller. These are shown to achieve different trade-offs between inherent complexity, requirements upon the system specification, flexibility and performance.

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1. INTRODUCTION

Fundamentally, a radar is a signal processing system. The distribution of antenna elements and sub-arrays defines a spatial sampling pattern; the individual elements apply a frequency and angle dependent complex weighting to signals; signals are then amplified, filtered and re-weighted, demodulated, adaptively combined and detected; the transmitted signal must be designed to match likely propagation conditions, to confuse countermeasures, and to achieve the best match to a particular type of target; the receiver must be matched to the expected return signal for optimum detection. Thus, the signal processing function is now seen as central to radar systems, involving the design of the waveform, antenna and information processing algorithms. This is in contrast with the traditional view which distinguishes between the "system" and the "signal processing", the latter being thought of as a back-end "black box" component. Conventional radars are analogue signal processors, whilst the current trend is towards the increased use of digital technology.

Although the signal processing required for a radar may be broken down into small blocks, the functionality of each block must be designed with the requirements of others in mind. Evaluating the choices available, and the consequent inter-relationships between such blocks, becomes increasingly complex when one considers multi-channel arrays. The decision to adopt adaptive nulling may affect the structuring of the antenna itself (design of subarrays, for example) the required digital wordlength in the ADC or output power, or may necessitate different ordering or partial duplication of other functions such as clutter rejection. Conversely, the choice of adaptive nulling scheme will be influenced by the practical constraints imposed by a given application. Whilst we have much experience of thinking about the structure of conventional radar signal processing chains, we have rather less knowledge concerning the integration of modern techniques such as adaptive nulling into multi-element systems.

This paper will present a review of the implications of alternative adaptive nulling architectures on the design of a space-based look-down radar, based on typical practical considerations. We will review the major constraints imposed by a space-based application in section 2, and summarise the beamforming requirements of such a system in section 3. Section 4 will define the sidelobe canceller and fully adaptive sub-arrayed nulling approaches and make further assumptions about the implications for hardware. In section 5 we will compare these alternative nulling schemes in more detail, taking as a basis the way in which they influence or are influenced by the partitioning of the phased array, receiver channel characteristics, ADC requirements, and other signal processing issues such as number of digital operations and convergence rates. Finally, in section 6, we will summarise the major advantages and disadvantages of each adaptive scheme for the space-based radar.

2. REVIEW OF SPACE-BASED RADAR CONSTRAINTS

In order to draw any useful conclusions, we must make some basic assumptions about the required radar specifications and the nature of the jamming threat and clutter environment which it will be required to deal with. These assumptions are made for the purpose of illustration, and are not meant to correspond to any operational design or requirement. Below, we summarise the detail given in the Appendix.

The radar will combine search and tracking functions, interleaved according to some radar control algorithm. Following detection, immediate look-back would be useful to confirm and to initiate tracking. A phased array will be necessary to achieve the beam agility required for such a system. We will assume a two-dimensional L-band (26cm) array containing 16384 elements (a convenient power of two). We will chose the array to be 28mx9m in size, giving an area of $24dBm^2$, azimuth and elevation beamwidths will be approximately 1.7° and 0.5°, respectively. A rectangular array, with its longer dimension aligned to the direction of travel, is chosen to ameliorate ground clutter returns. Peak total power will be 100kW, and the transmission will be of a range-ambiguous pulse-Doppler waveform. We assume a few 10ms coherent bursts before the beam moves on to a new position. We will chose signal bandwidth to be 1MHz, varying over a 200MHz agile bandwidth.

As can be seen from the detailed calculations in the Appendix, we will assume a range of possible target signal-to-noise ratio (SNR) between -98dB and -69dB at each antenna element, depending on the angle of scan, the size of the target, and losses. We estimate the likely main beam clutter to noise ratio (CNR) at the element to lie between -39dB and -13dB. Powerful ground-based jammers will be able to transmit into the antenna sidelobes over the full agile bandwidth, giving a jamming to noise ratio (JNR) of around 59dB. Less powerful airborne jammers may be able to transmit into the main beam over the full agile bandwidth, resulting in a maximum JNR of around 4dB at the element.

3. BEAMFORMING REQUIREMENTS

If we consider the integration of adaptive nulling with otherwise fairly "conventional" signal processing, we will need a sum beam and two difference beams for two-axis monopulse, and sidelobe blanking channels. Because of the nature of the system, ground clutter returns of significant bandwidth will be present in the data, and some form of clutter cancellation scheme will be necessary. This will not be considered in detail in the present memorandum. However, if we consider using the displaced phase centre antenna (DPCA) [1] technique for clutter removal, then all the beamforming hardware (apart from the second difference beam, which would no longer be required) must be duplicated for each of the two (or more) DPCA channels. We must maintain a high gain mainlobe in the sum beams for SNR enhancement (42dB with sixteen thousand elements), and a notch at boresight in the difference beams for accurate angle estimation. A random sidelobe floor of -15dBi must be achieved, partly for the suppression of sidelobe clutter, but particularly for the suppression of sidelobe jamming, which even then will have to be adaptively nulled by a further 50dB or more in each beam. Mainlobe jamming must be nulled by a similar amount.

We will assume for the purpose of subsequent analysis that 32 degrees of freedom will be sufficient to cope with the likely jamming threat. This choice will influence the settling time of the adaptive process, the peak sidelobe level (which must be minimised) and the stability of the beam pattern. A stable beam pattern is necessary to avoid modulating clutter. A smaller number of adaptive channels would be suitable if only a limited number of jammers are likely to be encountered. This would lead to faster convergence of the adaptive algorithm. On the other hand, a larger number of channels would also be technologically feasible.

4. ADAPTIVE CANCELLATION SCHEMES

We consider adaptive antenna schemes to fall into three broad categories. In each case, the phased array will be divided into sub-arrays in order to reduce the number of degrees of freedom from 16000 to a more reasonable number (taken to be 32 for this study).

4.1. SIDELOBE CANCELLER (SLC)

The sidelobe canceller makes use of one or more directional "main beams" and a number of less directional, lower gain "auxiliary beams". The main sum and difference beams would be formed by analogue beamformers, making use of the majority of elements in the array. The remaining elements may be taken singly or in smaller sub-arrays to form the auxiliary beams. Each element of a main beam will be subject to a fixed analogue beam-tapering weight, which is designed to lower the "natural" sidelobes of the beam [2]. In general, the auxiliary channels may or may not have zero gain on boresight, although in our system they need zero gain for the difference beam. Adaptive processing will change the unadapted beam shapes. Nulling signals which are in the main beam region could cause significant skewing of the main beam pattern. The maintenance of high sum-beam and zero difference-beam gain on boresight, following adaptive processing, relies on the auxiliaries having much lower allowable gain than the main beam. This also prevents the effective nulling of jamming signals entering the main beam and thus minimises the likely distortion of that beam.

As shown in Fig. 1, associated with each main and auxiliary beam is a receiver channel, containing

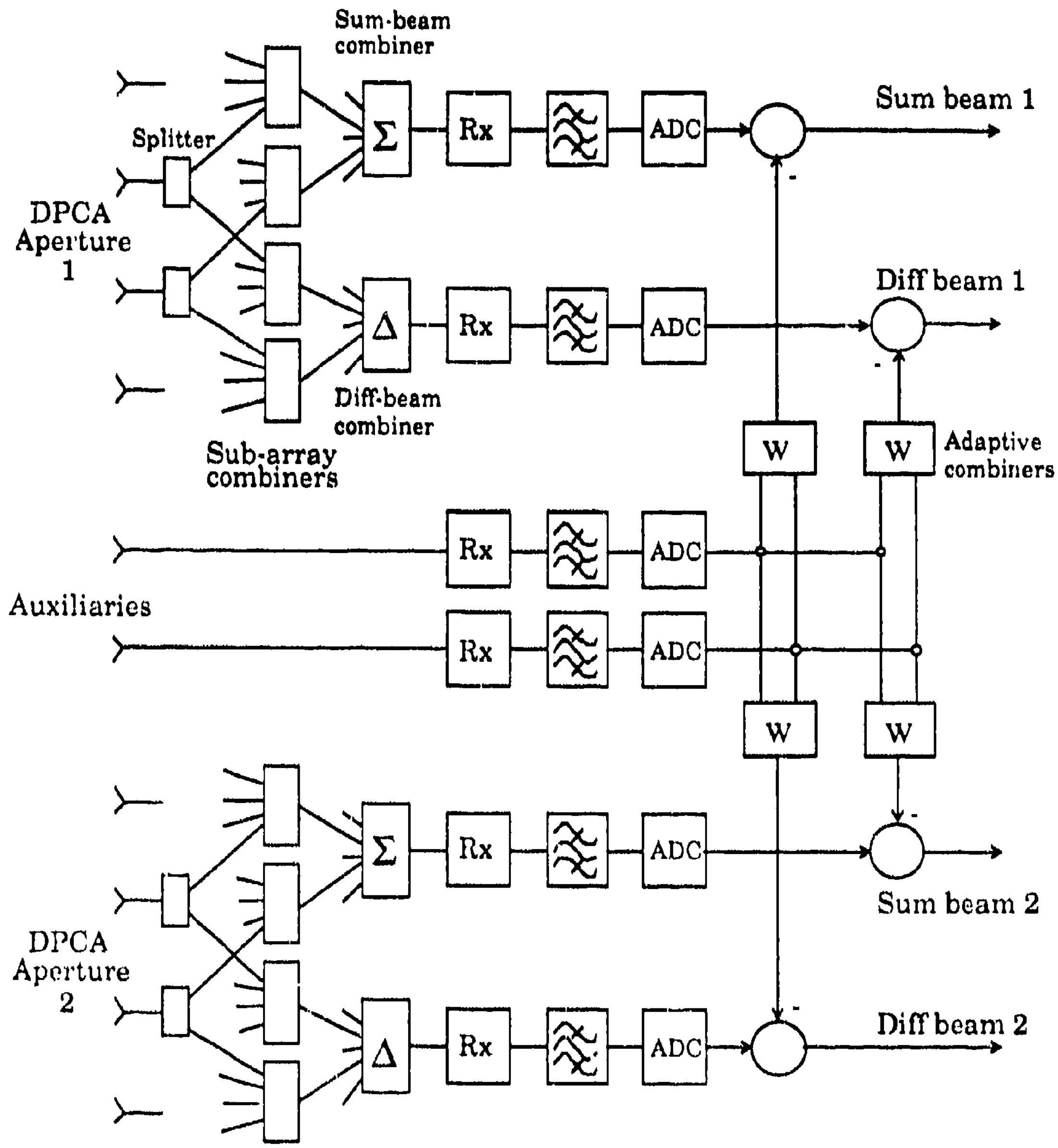


Fig. 1. Architecture of adaptive sidelobe canceller. The two main apertures are required for DPCA clutter processing.

signal-bandwidth-defining filters and ADCs. The adaptive algorithm will be implemented digitally. DPCA requires an extra, displaced main aperture and therefore the duplication of each main receiver channel. Although they are shown separated for clarity, the two main apertures would overlap considerably in practice, most elements contributing to both with their outputs split four ways via programmable attenuators. The auxiliary channels do not need to be duplicated: they can be adaptively combined with signals from both main apertures for sidelobe cancellation.

DPCA processing, although not shown, is assumed to follow adaptive beamforming, but there is no obvious reason why these two processes should not be reversed. It will certainly be necessary for clutter to be removed from the data which is to be used for the adaptive weight computation. This

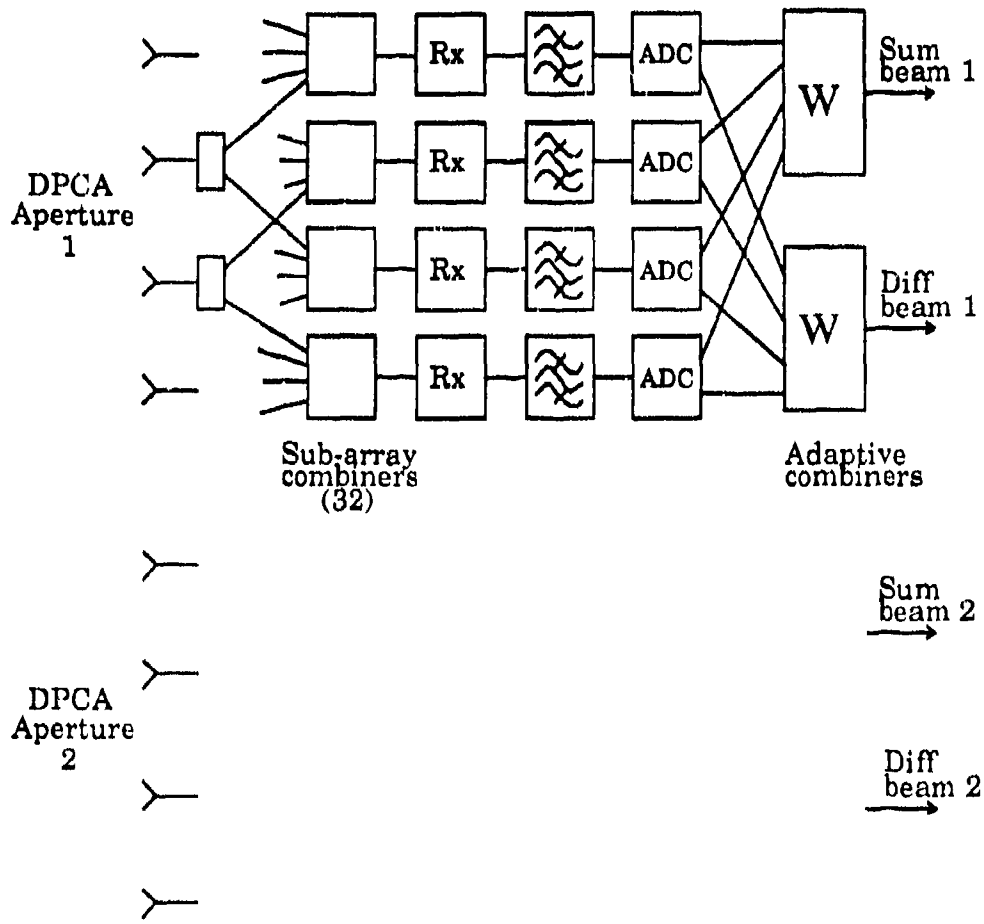


Fig. 2. Architecture for a fully adaptive subarrayed antenna. Each subarray is treated equally, apart from any overall aperture weighting function. The two main apertures are required for DPCA clutter processing.

will prevent adaptive degrees of freedom being used for clutter cancellation, and maximise those available to remove jamming.

4.2. FULLY ADAPTIVE ARRAY (FAA)

It is clearly impractical and unnecessary to make use of sixteen thousand degrees of freedom, and so a true fully adaptive array is not being considered here. Instead we mean an array which has been divided into a number of equivalent subarrays, each of which contributes a degree of freedom. These degrees of freedom are used either in the adaptive processing, or to control the beamshape through programmable constraints.

As shown in Fig. 2, the signal from each element of the array would be split and subjected to analogue beam-tapering weights before being fed to two different sub-array outputs, one that is optimised for contributing to a sum beam and the other to a difference beam. Each sub-array channel would then have its own receiver channel, containing downconverters, signal-bandwidth-defining filters and ADCs. The adaptive algorithm will be implemented digitally, although the adapted weights may be either digital or analogue. Each main beam is formed through the action of a directional constraint on the adaptive weight calculation. Normalisation of the weights with respect to the desired signal level at the output will cause the "gain" in the desired signal direction to be fixed. This does not imply that the output signal to noise ratio of an adapted system will always be the same as that which can be achieved by a fixed-weight system which is not subjected to jamming. For example, the suppression of main-beam jamming will cause an increase in the norm of the weight vector in order to meet the desired output signal level, and this will result in amplification of the system noise.

The inclusion of DPCA does not necessarily mean that the number of subarray channels must be doubled. That number was fixed at 32 for adaptivity, not for beamforming purposes. It might therefore be possible to redistribute the subarrays so that four sets of eight are each optimised for contributing to either a sum or difference beam at one of two phase centres. As in the SLC, the outputs from most elements would therefore be split four ways via programmable attenuators. Also as discussed for the SLC, it will be necessary to remove clutter from the data used by the adaptive weight computer. This is not shown in the diagram, and, although important, methods for clutter removal are not explicitly considered in this memorandum.

4.3. GENERALISED SIDELOBE CANCELLER (GSLC)

In this configuration, all antenna elements will be used to form the main sum and difference beams. The *same* elements, grouped into smaller sub-arrays, will provide auxiliaries for adaptive nulling. Two sets of beam-tapering weights will be required to shape the main and sub-array beams. Ideally, the auxiliary beams will have zero gain in the main beam direction, in order to maintain gain towards the desired signal following adaptation. It is likely, using current technology, that both the main beams and auxiliaries will be formed using analogue beamformers because of the number of elements involved. The adaptive algorithm is again implemented digitally, and it may be possible to apply additional constraints at this point.

The generalised sidelobe canceller of Griffiths and Jim [3] falls within this class of adaptive systems. This system makes use of all elements for both the main and auxiliary channels. The auxiliary channels are arranged to have zero gain in the main-beam direction. If the auxiliary beams are orthonormal to the main beam, and all degrees of freedom are used, then the system will be exactly equivalent, in principle, to a fully adaptive array having the same sub-array pattern. Mathematically, each is a linear transformation of the other. An alternative linear transformation of the fully adaptive array, proposed by McWhirter [4], leads to another generalised sidelobe canceller, which van Veen [5] has shown to be equivalent to a factored implementation of the Griffiths architecture.

5. COMPARISON OF NULLING SCHEMES

As shown by the figures in the Appendix, in all cases under consideration, the desired signal will be below the noise level at the input to the adaptive nulling computation. This may be important if the same data is used to calculate the adaptive weights. This is because the presence of significant signal in or around the desired main beam direction causes some deterioration of the attainable output signal to noise ratio. As we shall see in section 5.2, the FAA has the capability to form nulls in the main beam and would be most sensitive in this respect. For the same reasons, main beam clutter must also be filtered below the noise level if it is not to capture adaptive degrees of freedom. In the Appendix clutter is shown as being above the noise level at the input to the ADC. For the purpose of comparing the nulling schemes, we will assume that it can be suppressed (perhaps using DPCA) prior to computing the adaptive weight vector.

5.1. PARTITIONING OF THE PHASED ARRAY

5.1.1. DEGREES OF FREEDOM

For greatest efficiency, in a two-dimensional phased array, sub-arrays must be positioned to give the maximum number of degrees of freedom in all planes normal to the array. For example, if a sixteen phase-centre array is arranged on a regular 4x4 grid, then only four degrees of freedom will be available in the principal planes. Flam [6] has shown how this effect may be simply understood, and concludes that a planar array consisting of 2M carefully placed phase centres should be able to form a minimum of M nulls in a given conical region of interest. In the MESAR phased array [7], a sub-arrayed system, the phase centre spacings have been randomised for this reason. Similar design rules will apply to the phase centres of a sidelobe canceller, in which it is likely that the auxiliaries will be distributed around the periphery of the main array.

5.1.2. GRATING LOBES

Sub-array placement will also be important from the point of view of grating lobes. These are images of the main beam, pointing in other directions. They result, in general, from having distances between adjacent sub-array phase centres of greater than half of a wavelength- that is, spatial under-sampling. The presence or absence of grating lobes will be governed by the separation of the phase centres as projected onto the plane of interest. It may be possible to extend the concept of minimum redundancy arrays [8] to the design of planar structures with widely separated phase centres. Grating lobes will cause problems with element-space adaptive systems such as the sub-arrayed FAA or GSLC, even though they may have been reduced to a level similar to the normal antenna sidelobes. Since a grating lobe is an image of the main lobe, in a FAA system its behaviour will be governed by the directional gain constraint applied to the main beam. If a jamming signal is directed at such a lobe, the system will be unable to adapt against it effectively. This problem will not occur with the SLC, unless individual antenna elements are too widely spaced. However, in either system, the creation of a null in the direction of a jammer may be accompanied by the formation of grating nulls in other directions. It is possible that these could interact with nulls directed against other jammers. Randomisation of the phase centre positions will suppress both grating lobe and grating null effects.

5.1.3. DISPERSION

Phase dispersion across the aperture, associated with wide-band signals located away from boresight, will increase the number of degrees of freedom required to adapt against a single jammer and may reduce the depth of the resultant null. These degrees of freedom may be spatial (occupying additional phase centres) or temporal (requiring adaptive FIR filters in each adaptive channel). Any increase in the number of degrees of freedom will cause an increase in the amount of hardware required, the digital computation load, and the time taken for the adaptive algorithm to converge and stabilise.

It has been suggested by Barton [9] that the SLC may be more sensitive to aperture dispersion effects than the FAA because of the large distances between phase centres. Since the overall apertures of the SLC and FAA systems which we are considering will be the same, we have no reason to believe that the sensitivity to aperture dispersion would be any different in each of the two cases. However, Barton points out that the delay which ought to be considered must also include the antenna feed networks. In some systems, the SLC main array may have a longer feed than that for the auxiliaries, causing the total delay to be much greater than implied by the physical aperture. Compensation will be necessary, perhaps by introducing additional line lengths into the auxiliary channels. The feeds associated with the sub-arrays of the FAA would naturally have better-matched dispersive properties. More importantly perhaps, the complex patterns of the SLC main and auxiliary arrays will be different and will vary differently as a function of frequency, whereas the patterns of the FAA sub-arrays should match much more closely. This effect therefore may increase the number of degrees of freedom required to null a given number of jammers more significantly with the SLC than with the FAA.

5.1.4. RELIABILITY

Finally in this section, it is perhaps worth referring to reliability. This is a complex issue, which we do not propose to consider in any detail. However, it is clear from a simple example that the different architectures implied by the different adaptive nulling schemes may have different reliability characteristics. If the main beam summation node in the SLC or GSLC should fail, then the system will cease to operate. If a sub-array summing node fails in the FAA, then we would only lose a single degree of freedom, in principle. However, this is a very specific failure, and appropriate hardware design could insure against it. An alternative view might be that the greater hardware complexity of the FAA would lead to a higher probability of some type of failure. Another important point is that failure of receiver channels during the lifetime of the system will have potentially serious consequences for achievable random sidelobe levels. Although the adaptive control of the sidelobe pattern will work to minimise the impact of this, the dynamic range required of the ADCs may have to be increased to take it into account (see section 5.2.2 for discussion of dynamic range).

5.2. RECEIVER CHANNEL CHARACTERISTICS AND ADC REQUIREMENTS

5.2.1. CHANNEL MATCHING

For optimal performance of the adaptive nulling system, it is important that the receiver channels should be well matched over the operating frequency band, for the same reasons as discussed in connection with dispersion in the previous section. The match should be maintained as a function of signal amplitude and operating temperature, the latter implying a need for careful temperature stabilisation in the severe conditions of space. Taking into account the antenna feeds and RF components, matching may pose a problem in a sidelobe canceller system. However, the receiver channels for the FAA degrees of freedom are nominally the same as each other, and one might expect that this would simplify the matching problem. Nevertheless, matching the channels of a FAA to the accuracy required for effective suppression of powerful ground-based jamming may still be very difficult. In order to achieve null depths of 50 to 60dB, matching equal to 0.1° rms in phase and 0.015dB rms in amplitude is likely to be necessary. Such a degree of match is also likely to be necessary for effective DPCA clutter processing.

A major source of mismatch in receiver channels is likely to be the signal-band-defining filter. Each channel contains such an analogue filter, and whilst it may be possible to match the filters at band-centre, it will be significantly more difficult to match the phase characteristics in the band-pass skirts. Although this mismatch may be minimised with careful design, it is likely that further action will be necessary. One possibility is to calibrate each channel at a number of spot frequencies across the band, and then to correct for differences using a digital FIR filter following the ADC. If the filter is re-programmable, then this would potentially allow in-service re-calibration. The match would now be limited by the accuracy of the calibration and the resolution of the filter coefficients. Another approach, which Pohlig [10] has shown to reduce mismatch problems, is to adaptively control a set of analogue weights which precede the filters. This technique could be used as well as or instead of the FIR filter described above, and could also be beneficial in reducing dynamic range requirements, as we shall see later in this section.

Another potential source of mismatch is in the ADCs themselves. For example, these may have nonlinear operating characteristics, and will have to have carefully matched aperture times. In the SLC, we shall see that there is apparently potential for reducing the system cost by using ten-bit converters in the auxiliary channels, whilst using a twelve-bit converter in the main channel. However, apart from the additional cost of manufacturing two different space-qualified ADCs, such a scheme may introduce additional tracking problems associated with the inevitably different characteristics of the two devices. The different quantisation steps of the two converters may also cause problems. An alternative approach, which is likely to be possible in future, would be to use "single-bit" oversampling ("sigma-delta") converters. This would reduce the difficulties of producing a linear conversion characteristic. Such converters typically sample the analogue signal at many times the normal rate, to produce a pulse-density modulated stream of bits, which are then digitally integrated to provide the required output digital wordlength. Converters based on similar principles, but using initial coding into three- or four-bit words in order to reduce the sample rate required, may be available for radar applications in the shorter-term.

5.2.2. DYNAMIC RANGE

For a FAA, the dynamic range requirement will be the same for each channel, whereas it will be different for the main and auxiliary channels of the SLC. The SLC, expected to cope principally with jammers in the sidelobes, will already have achieved a significant degree of jammer suppression through the tapered sidelobes of the main beam and the correspondingly lower auxiliary channel gains. Thus, the wordlength requirements of ADCs should be eased, as will those of the adaptive cancellation computation.

The calculations given in the Appendix derive possible dynamic ranges of signals, as measured at the antenna elements. By following these signals through the different architectures, as shown in

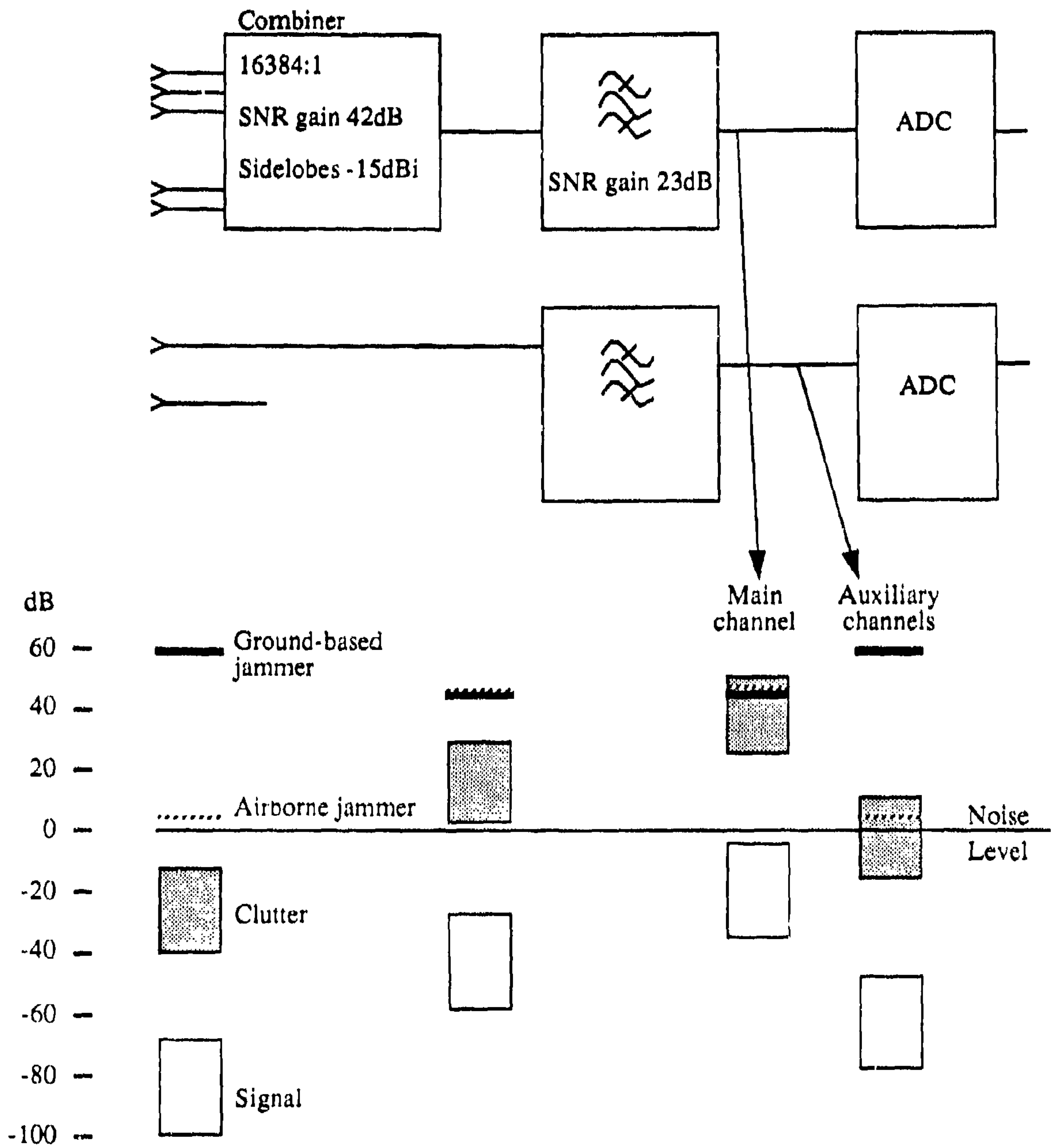
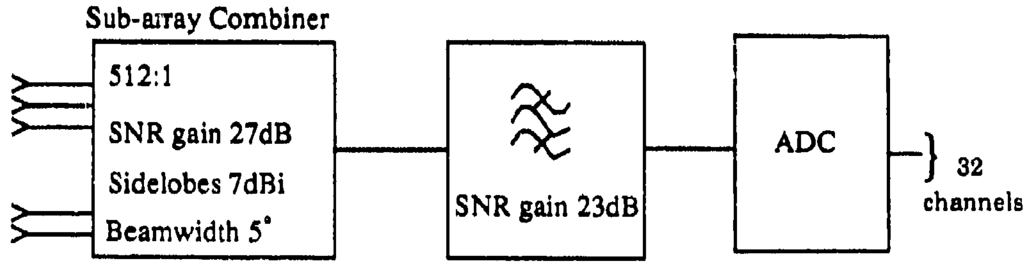


Fig. 3. Signal, clutter and jamming levels with respect to noise, in an adaptive sidelobe canceller array, such as that shown in Fig. 1. The upper chain corresponds to the main channel, and the lower chain to one of the auxiliaries. The powers shown at element level correspond to those given in the Appendix.

Figs. 3 and 4, we see how the dynamic range changes as a result of subsequent processing. For example, in either the SLC or FAA systems, at the element level we find a possible worst-case target signal-to-noise ratio of around -98dB, and a worst-case jammer-to-noise ratio of around 59dB. As we progress through each system, we find that the signal, jammer and clutter levels are changed in different ways. For simplicity in the diagrams, we consider systems that form a sum beam only.



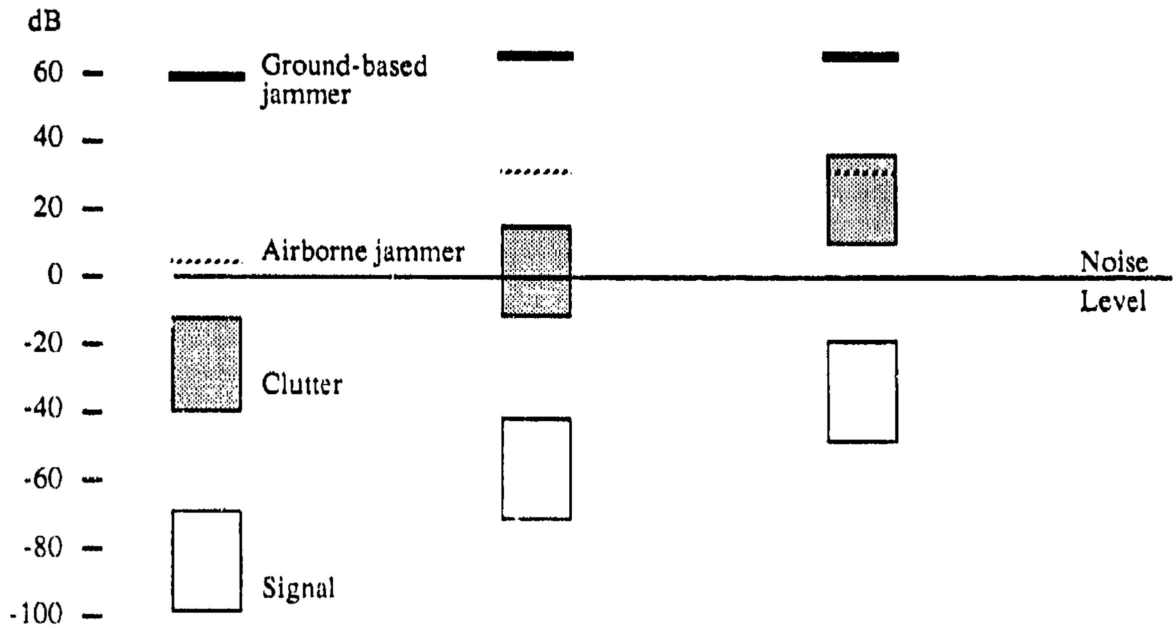


Fig. 4. Signal, clutter and jamming levels with respect to noise, in a subarray channel of a fully adaptive array, such as that shown in Fig. 2. The powers shown at element level correspond to those given in the Appendix.

In the main sum-beam channel of the SLC, the gain of the mainlobe raises the minimum SNR to -56dB and the airborne JNR to 46dB, whilst the -15dBi tapered sidelobes reduce the ground-based-JNR to a similar level. The signal-band-defining filters cause a further increase in SNR to a minimum of -33dB in the main channel. Thus the dynamic range, defined by noise at one extreme and the maximum jammer or clutter level at the other, would probably allow us to work with a 12 bit ADC, whilst still encoding sufficient information concerning the "noise plus signal" statistics to enable the system to integrate and recover the signal at a later stage of the processing. In addition, although the dynamic range requirement of the auxiliary channels would *seem* to be even greater than that of the main channel, it can in fact be lower. Since the only purpose of the auxiliaries is to pick up sidelobe jamming, their sensitivity can be reduced, using either an AGC or a fixed attenuator, in which case only 10 bits may be sufficient in principle. This will reduce the level of airborne main-lobe jamming signals fed to the adaptive processor, and so reduce the capability to null these signals. In an SLC system, this type of main-lobe jamming is likely to be difficult to deal with in any case, because of the very high weight norm required to achieve sufficient null depth. Indeed, we have already accepted the inability to reject such signals by assuming a constraint on the norm of the adaptive weight vector. In the presence of such main beam jamming, the SLC radar would simply have to look elsewhere, and we must accept that this may enable an airborne jammer to deny important coverage to the radar. However, as mentioned in the previous section, using different ADCs in this way may exacerbate the problems of channel matching, and use of an attenuator of AGC is likely to offer a more suitable solution.

In order to reduce the dynamic range needed by the SLC main channel to deal with sidelobe jamming, Pohlig [10] has proposed a hybrid analogue and digital nulling scheme in which some cancellation is achieved before digitisation. This is achieved without loss of SNR. The weight computer produces a set of analogue weights which are fed back to the auxiliary channels at a point preceding the analogue filters. Digital weights are fed forward and are found to improve the convergence characteristics of the technique. Ward et al [11] have also described a related closed-loop feedback architecture. In both cases, we assume that the (digital) computation of the analogue weights will converge, given a starting point which may include signals which have been clipped by ADC overflow.

By contrast with the SLC, we see that the assumptions made concerning the characteristics of the sub-arrayed FAA lead us to a requirement for a 14 bit ADC in each of the 32 channels. This occurs because the antenna beam pattern for the whole aperture is not formed until after adaption, and so we do not have the pre-formed low sidelobe pattern of the SLC with which to gain some immunity from jamming. Since it has not been amplified by a full main-lobe gain, the situation for airborne jamming might appear to be easier than with the SLC. However, it would still be difficult to null such jamming adequately since attenuating by 34dB, instead of providing 15dB of gain, as the final stage of beamforming should, is equivalent to 49dB of cancellation. At the very least, this would cause severe distortion of the main beam and (more importantly and fundamentally) a significant increase in the noise level following adaptation, due to the increase in the weight norm needed to create such a null. However, the system would still have maximised the output signal to noise plus interference ratio, as required by the adaptive weight computation, and so the output would be an improvement over the quiescent un-adapted case.

The estimates of ADC dynamic range, discussed above, may have to be further modified to take account of the likely increase in sidelobe levels resulting from channel failure and degradation over the lifetime of the array. Yet another influence on the choice of digital wordlength may be the likelihood of higher peak jammer powers caused by constructive interference of multiple jammers. To some extent, both of these effects are allowed for in the estimates given above, but would require further analysis for a specific design.

5.2.3. TAPERING

The ability to accurately "taper" the aperture, as described in section 4, to achieve low sidelobes implies accurate knowledge of the channel calibration and match. In the case of the SLC, aperture tapering is clearly important in that it effectively partitions the task of sidelobe jammer suppression between a fixed analogue section and an adaptive digital section. This reduces the dynamic range of the receiver channels and the ADCs, and consequently reduces the wordlength needed in subsequent digital processing. As seen in the previous section, this benefit is not obtained in an FAA, even though tapering is applied at the elements. Nevertheless, Gupta and Ksienski [12] have shown how tapering is equally important for the fully adaptive array. They have demonstrated that the output signal to noise plus interference ratio (SNIR) depends on the conventional un-adapted beampattern. Therefore, lower quiescent sidelobes can result in a higher adapted output SNIR. However, for all the adaptive schemes under consideration it is worth remembering that aperture tapering to gain improved sidelobes will also lead to an increase in the main beam width. Since the SLC is unable to null jamming within the main beam, this will be a limitation and should be taken into account in deciding the physical aperture. Similar comments apply to the FAA, as nulling main beam jamming can lead to a lower output SNIR than for sidelobe jammers, as described above.

5.2.4. BANDWIDTH

This issue is related to the discussion on dispersion effects. To some extent, wide band jamming may be nulled using spatial degrees of freedom, since the time taken for the signals to cross the aperture may provide the delay necessary for such processing. The number of degrees of freedom required will depend on the dispersion of the jamming signal across the aperture, which in turn depends on the bandwidth and angle of arrival of the jamming signal. Hudson [13] has examined how degrees of freedom are used as a function of bandwidth. If it is known *a priori* that jamming signals are likely to use up multiple spatial degrees of freedom, then adding further adaptive channels to the system could provide a straightforward solution.

Perhaps a more natural solution to the problem of significant jammer bandwidth would be to simultaneously adaptively adjust the combined spatial and frequency response of the channels. Given a signal requiring more than one degree of freedom to form a null, then there would be a choice of architectures for both the SLC and the FAA. One possibility would be to use an FFT to break up the signal bandwidth, followed by multiple "narrow band" adaptive canceller algorithms acting on frequency cells containing significant power. This may have convergence performance similar to the narrow band case, but there is an obvious hardware penalty. An alternative would be to carry out wide-band nulling using an adaptive tapped delay line architecture [14]. This is likely to converge more slowly because of the greater number of degrees of freedom. For the FAA and GSLC, constraints would need to be designed such that the frequency response of the adapted system is unperturbed in the direction of the desired signal.

5.3. OTHER PROCESSING ISSUES

5.3.1. NUMBER OF DIGITAL OPERATIONS

Open-loop computation of the adaptive weight vector typically demands $O(N^3)$ complex numerical operations, where N is the number of adaptive channels. Using combined parallel and pipelined (systolic) processing architectures, these operations can be carried out in $O(N)$ time steps. We have concluded in section 5.2 that the FAA will require longer digital wordlengths than the SLC in all channels. Therefore, the subsequent digital processing must also use longer wordlengths. This implies additional bit-level operations and may slow down the maximum processing rate. However, as we shall advocate the use of parallel or pipelined processors, possibly working in the data domain, this is unlikely to be very significant. As the power of standard DSP chips continues to increase, processing load is no longer thought to be a dominant issue in real-time system design. As Knowles et al have shown [15], even recursive operations may be carried out at a rate which is independent of wordlength.

5.3.2. CONSTRAINTS AND ROBUSTNESS

The shape of the main beam of the SLC will remain relatively undistorted following adaptation. This is because nulling of mainlobe jamming will be prevented, either by reducing the gain of the auxiliaries to such jamming, or by constraining the maximum adaptive weight norm (which may restrict the choice of adaptive algorithms in a digital processor). Since the auxiliary channel gains, including weighting, will be therefore much less than the gain of the main beam, mainlobe perturbations will be small. Thus, the shape of that beam is principally determined by the physical design, electromagnetic properties and the analogue aperture tapering weights. The more accurately the antenna may be characterised, the more accurately the beamshape and main channel sidelobe levels may be predicted.

The main beam of the FAA is created after the application of adaptive weights. The gain in the desired pointing direction is determined by a linear constraint on the calculation of the weight vector. Under quiescent conditions (in the absence of jamming), knowledge of the resultant beamshape will again depend on how accurately the antenna and receiver system has been calibrated, and on the analogue aperture tapering function. Whilst adaption against sidelobe jamming will not greatly perturb this shape, main beam jamming may lead to significant distortion of the beam. This would appear to have dire consequences for monopulse direction-finding. However, in principle, the distortion can be predicted and corrected for [16]. The extent to which this can be achieved depends on the proximity of the jammer to the constraint direction and on the JNR. The closer the jammer is to the direction specified by the constraint, the higher the weight norm required to null it, and the higher the dynamic range required of the digital part of the system. Although the beam distortion may be corrected, and monopulse direction finding capability restored, the increased weight norm will amplify the system noise. This means that the output SNIR will be lower than in the presence of a sidelobe jammer of the same power, and consequently the estimates of direction of arrival of the wanted signal will have greater variance.

If the desired signal is present above the system noise level in the data used to calculate the adaptive weights, the FAA becomes extremely sensitive to the accuracy of the beam-steering constraint. If the signal arrives from a direction different from that assumed, then the system will attempt to null it, sharply reducing the output SNIR as well as distorting the beam [12]. We have shown in the Appendix that the signal in the present application will be below the noise, and so this will not be a problem. If the desired main beam has the same shape as the SLC main beam then sensitivity to error in the pointing direction will be determined by the rate of change of gain around the nose of the beam. In general, this will be the same in each case, although beam skewing in the FAA may lead to higher rates of change under some circumstances. For this application, additional constraints on the FAA, such as derivative constraints or additional gain constraints, are unnecessary and would only serve to reduce output SNIR. However, further "soft" constraints [17] may be useful in order to "tie-up" otherwise un-utilised degrees of freedom, thus reducing the tendency towards instabilities in the sidelobe pattern (discussed further in the following sub-section). Soft constraints are added to the data such that the adapted sidelobe pattern tends towards that for the un-adapted system in regions not affected by jamming.

Sidelobe clutter has not been mentioned so far in this memorandum, nor as far as we are aware in the adaptive-nulling literature. It needs to be considered, however, because it will be above the thermal noise level in the SLC main channel and considerably more so in the FAA channels because of their inferior sidelobe suppression. It will have a very wide Doppler spread, and may not be removed efficiently by DPCA techniques because of the effects of mis-match between the two antenna patterns. If such clutter cannot be rejected independently from the data used for adaptation, it will capture the degrees of freedom remaining after suppression of the strongest jammers. In this event, its effect would be simply to lower the general sidelobe level of the adapted beam. Given the limited number of degrees of freedom in a partially-adaptive system, such as the SLC or the FAA being considered here, this would adversely affect the ability to reject weaker jamming signals, and would, in any case, have a limited effect on spatially distributed clutter. Furthermore, a single set of adaptive weights needs to be calculated and used for a given coherent integration period. If this is not done, a fluctuating sidelobe pattern will result from the varying sidelobe clutter, and any subsequent suppression of sidelobe prior to target detection will be adversely affected. This topic needs further examination.

5.3.3. CONVERGENCE RATES

It has been shown that an adaptive system using a direct algorithm for calculating the weight vector will converge to within 3dB of the optimum output SNIR within approximately 2N data snapshots, where N is the number of channels, in the absence of the desired signal [18]. This should be true for both the SLC and FAA systems in the presence of sidelobe jamming. However, because of the additional gain constraint employed in the SLC the two architectures have different behaviour where beam patterns are concerned.

In the SLC, the main beam and its sidelobes are formed in advance of the adaptive processing. The limitation of auxiliary channel gain, through a norm on the adaptive weight vector, results in an adapted pattern which has sidelobes within a few dB of the quiescent pattern. In the FAA, on the other hand, the basic beam pattern is not synthesised in advance and it is not usually subject to constraints on channel sensitivities or weight norm. Although nulls and main-beam are formed quickly, the sidelobe levels resulting from different weight vectors may vary (or "jitter") significantly. The achievement of uniformly low and stable sidelobe levels may require many more data snapshots to be used for the weight calculation. Sidelobe "jitter" might result from excess degrees of freedom being used to model correlation in the sidelobe-clutter, or random cross-correlations in the noise caused by the limited number of data snapshots. Even though this has little effect on SNR, it could lead to higher than desired sidelobe levels and consequently higher levels of received clutter. If weights are recalculated at shorter intervals than the coherent dwell time, the jitter may also affect subsequent clutter rejection processing, as noted earlier.

5.3.4. FLEXIBILITY

A FAA or GSLC is able to apply simultaneous multiple constraints at the pre-processing stage in order to control main beam gain and shape (through a steering constraint and derivative constraints) or to maintain gain in multiple directions. In either case, as the number of constraints increases, there will be a loss of output SNR with respect to the optimum. With multiple directional gain constraints applied as a pre-processing transform, convergence may also be adversely affected. Rather than fixing the response of a single beam pattern at many angles, multiple FAA beams may be simultaneously optimised in a number of different directions, without loss of output signal to noise ratio or convergence, by using a digital post-processor to apply the constraint. A single computation of the matrix inverse required by the adaptive computer is sufficient for the formation of beams covering a range of angles. This concept has been demonstrated in the McWhirter/Shepherd MVDR beamformer [19].

The FAA also offers the choice of working directly on the data (or covariance estimate), or processing in beam space. In principle there is no difference in the weights calculated by the two approaches. In practice, points to consider would include the effects of grating lobes, impact on dynamic range, and the flexibility to apply non-interacting constraints in a post-processor. For a system with a large number of degrees of freedom, transforming into beamspace may allow a reduction in the number of degrees of freedom ("rank reduction") through choosing to process only those beams with significant power. This approach would accelerate convergence and stabilise beam patterns, with only slight loss of output signal to noise ratio. Alternative approaches to beam stabilisation are the Brandwood [20] trapezoidal decomposition method, using the singular value decomposition to reduce the variance of the data matrix, or the Gabriel beamspace method based on high resolution pre-processing [21])

Few of these choices are available for the SLC. Therefore, the FAA seems more flexible. Whether this flexibility is sufficient to encourage the use of the FAA would depend on the mission requirements and a determination of the practical value of these options.

5.3.5. CHOICE OF ADAPTIVE ALGORITHM AND ARCHITECTURE

From the point of view of rapid convergence to a stable result, Sample Matrix Inversion (SMI) [18] operating on a covariance matrix derived from the measured data, or QR decomposition [22] operating directly on the data, seem to offer the best choices of algorithm. Both compute the adaptive weight vector in open-loop fashion. Both result in an output SNR within 3dB of the optimum when calculated from 2N+3 data snapshots (if signal and mainbeam clutter are excluded from these data). Faster algorithms exist (for example Hung-Turner [23]), but inevitably result in lower output SNR because of the poorer averaging of the noise and consequent poor estimation of the optimum weight vector.

The likelihood of high dynamic range in a FAA suggests that QR decomposition would be the favoured algorithm. This is because formation of the covariance matrix required by SMI requires an increase in the computational wordlength, whilst still only representing the same information. Thus, 14-bit data immediately suggests a minimum wordlength requirement of 28-bits for subsequent processing. QR decomposition can be applied directly to the data, and, since the subsequent processing principally involves orthonormal rotations of this data, wordlength does not increase significantly during the weight calculation. Even if SMI is chosen as the preferred algorithm, QR decomposition provides a numerically secure approach for the necessary triangularisation of the covariance matrix.

Computationally efficient systolic architectures exist for both narrow-band [3, 22] and wide-band [14] implementations of the QR algorithm. These processors break down the task of matrix decomposition into a set of simple elementary transformations which can be carried out in parallel and pipelined fashion. The efficient use of many devices carrying out such simple operations in parallel results in architectures which actually scale in power to match the size of the problem (increasing number of degrees of freedom). Given the rapidly increasing power of standard DSP chips, an alternative approach could be to farm out entire matrix decomposition onto individual microprocessors [24]. Each would calculate a weight vector appropriate to a different sample of data. Following an initial period of latency, updates could be provided in rapid succession if necessary. In future, alternative 64 bit floating point DSP microprocessors, using number theoretic transforms to obtain wordlength-independent processing rates may possibly make covariance-domain processing even more viable in this context. With any of these approaches, both on-line and off-line architectures are possible [10, 25]. However, it is likely that the off-line solution with periodic weight update would be used in practice, since this allows greater control over the update rate (to match the coherent integration period), and also enables calculations to be carried out on a sub-set of the received samples. This may imply the need for more systolic nodes in order to extract the required weight vectors in parallel, or a corresponding reduction in the achievable weight update rate. The SLC, with its apparent lower dynamic range requirement, may be able to make use of less numerically sophisticated algorithms and alternative processing architectures, such as in the FLAP processor used for MESAR [26].

A further alternative possible with the FAA would be to use a high resolution algorithm in place of the usual adaptive processing [15, 21]. (The broad coverage of the main beam of a SBR suggests that high resolution may be desirable in any case.) Such an algorithm would provide direct estimates of jammer parameters, such as direction of arrival, power and cross-correlation. The estimated jammer positions output from such an algorithm could be used to form "deterministic" nulls. By this we mean that steering vectors corresponding to estimated directions of jamming sources would be used in place of data in order to construct the appropriate weighting vector. This would also confer stability when processing in the presence of a strong desired signal. Alternatively, if the desired signal is below the noise level as we have assumed, then the weight vector output by the high resolution technique itself may be used to null jamming. It may also be possible to carry out "high resolution" processing following pulse compression and coherent integration, using techniques such as IMP [27,28] or PTMF [29]. This could provide a way of handling the problem of estimating the direction of arrival of the main signal. Beam skew, caused by nulling of main-beam jamming, would no longer be apparent as a problem.

6. SUMMARY

In the context of an imagined space-based phased array radar, we have examined the application of three different adaptive nulling schemes. We have shown that the design of the "front-end" signal processing (from antenna design through to ADC design and clutter processing) has an impact on the use of an adaptive algorithm. Conversely, the decision to make use of adaptive nulling places constraints and strict performance requirements on other parts of the system. The "system" clearly cannot be distinguished from the "signal processing". We see, not surprisingly, that the radar must be designed as a whole, taking into account trade-offs and interactions between components.

In addition, our review has led to a number of interesting conclusions regarding the suitability of the different adaptive nulling architectures to the space-based application. These are summarised below, and are also given in the table.

Fully adaptive arrays would seem to have a number of broad advantages:

1. Effective main beam nulling capability, but accompanied with distortion of the un-normalised beamshape and lower output SNIR than for sidelobe jamming.
2. Greater flexibility through the use of programmable constraints, the ability to simultaneously make available more than one optimised output, and the option of carrying out high resolution processing.
3. Potentially lower sensitivity to aperture dispersion effects.
4. All channels (nominally) the same. The FAA sub-array beam patterns will all have similar frequency sensitivity and are more likely to track as a function of amplitude, time and temperature.

Table: Summary of advantages and disadvantages of alternative architectures

	SLC	FAA	GSLC
Hardware	Simplest/cheapest	Higher hardware complexity	Varies according to detailed implementation
Degrees of freedom	May need more to cope with dispersion across large main aperture	May need more for beam-shape constraints	"
Grating lobes	None, but will have grating nulls Must suppress via random sub-array placement	Must suppress via random sub-array placement	"
Dispersion	May be slightly greater due to mis-match of main and auxiliary beam tracking with frequency	Sub-array beams and channels should have similar characteristics	"
Channel match	Same sensitivities as others - band defining filters and ADCs	Same sensitivities as others	Same sensitivities as others
Dynamic range	Significantly lower than FAA - 12 bits should be adequate for all channels	High - will require 13 or 14 bit ADCs in each channel	Varies according to detailed implementation; probably between SLC and FAA
Operations	Needs shorter wordlength	13-14 bit data implies longer word-length in adaptive computer	"
Flexibility	Zero without substantial extra hardware	High - could use super-resolution, or multiple look directions	Probably could use additional programmable constraints, other-wise low or at cost of hardware
Convergence	Faster settling of sidelobes than for FAA	Weight jitter results in high side-lobes which converge quite slowly	Somewhere between SLC and FAA, depending on detailed implementation

SLCs might have the following advantages:

1. Lower hardware complexity and therefore lower cost.
2. Greater robustness of the main beam shape in the presence of close-in jamming, enabling straightforward monopulse direction-finding (and, conversely, less ability to cope with main-beam jamming).
3. Better control over mean sidelobe levels - the SLC, as a result of the constrained auxiliary gains, seems to be less dependent on convergence of the adaptive algorithm in order to maintain low sidelobes.

4. A lower dynamic range is required of the adaptive circuits if powerful jamming enters only via (attenuated) sidelobes;
5. It is easier to avoid problems with grating lobes - the system would not be asked to form nulls in directions equivalent to the constrained main beam.

It is difficult to draw such simplified conclusions about the broad category of "generalised" sidelobe cancellers, as these may have a range of features which overlap the two extreme types, and which depend on the detail of the particular design. Clearly, those systems which are mathematically equivalent to the FAA should have broadly the same advantages and disadvantages (although it is difficult to see how post-processing constraints would apply). If the system is restricted to sidelobe cancellation, then it may be designed to benefit from the lower dynamic range, as with the SLC.

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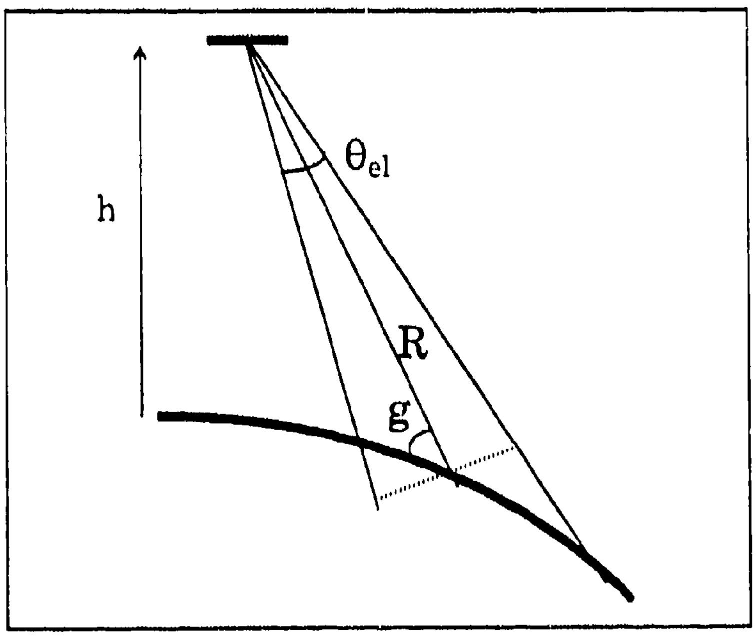
APPENDIX

DETERMINATION OF SIGNAL-, CLUTTER-, AND JAMMER-TO-NOISE RATIOS AT ANTENNA ELEMENTS

Assume the following radar parameters:

Wavelength λ = 26 cm
Antenna area A = 28 × 9m
Beamwidths θ_{az} = 1.7°, θ_{el} = 0.5°
Number of antenna elements 16384 (i.e. 2^{14})

Peak power P = 100kW
Duty ratio = 10%
Pulse-repetition frequency = 10kHz
Agile bandwidth, Ba = 200MHz
Signal bandwidth = 1 MHz
Target cross-section σ_t = 1 - 10 sq. m.
Noise figure F = 3dB
Losses L = 10 - 13dB + scanning loss where appropriate



Height, h = 1000 km
Scan angle = 29 - 58°
Taking the radius of the earth to be 6400km,
Slant range R = 1172 - 2665 km,
Grazing angle, g = 56 - 11.3°.

For SNR and CNR, best and worst cases will be considered in order to obtain maximum and minimum values. By "best" we mean best from SNR point of view, i.e. 29° scan angle. This angle also produces the highest CNR.

(i) SIGNAL TO NOISE RATIO (SNR)

Peak SNR at element per pulse (i.e. before integration) = $\dfrac{PA\sigma_t}{16\pi R^4 LkTFBa}$

	Best case		Worst case	
	+dB	-dB	+dB	-dB
PA	74		74	
σ_t	10		0	
16π		17		17
R^4		243		257
L		11		16
kTF	201		201	
Ba		83		83
	....	....	....	....
	+285	-354	+275	-373
SNR	-69dB		-98dB	

(ii) CLUTTER TO NOISE RATIO (CNR)

If the effective cross-section σ_c of the ground covered by one range cell can be determined, it can be substituted for σ_t in the above expression to evaluate mainlobe CNR. This is most easily, although approximately, done by first considering the dotted line in the diagram.

The dotted line represents approximately the length of Earth's surface illuminated by the beam, projected onto a plane perpendicular to the direction of the beam. It has length $R\theta_{el}$. The length of one range cell and its ambiguities is then $R\theta_{el}$ x duty ratio. (The effect of beam broadening on θ_{el} due to scanning of the beam may be disregarded, because scanning loss will cancel it out.) The area of a range cell is therefore $R^2\theta_{az}\theta_{el}$ x duty ratio.

The effective echoing area of each sq. m. of projected area is γ, the normalised backscattering coefficient. We assume γ lies between -10 and -20dB sq. m. per sq. m., and is constant with grazing angle. The total effective echoing area is therefore

$$\sigma_c = R^2\theta_{az}\theta_{el} \cdot \gamma \cdot \text{duty ratio . beam-weighting factor.}$$

The beam-weighting factor is to account for the reduction in width of ambiguous range cells towards the edge of the beam, and also for the reduction in transmit power reaching these cells.

	High grazing angle	Low grazing angle
γ	-20 to -10dB	-20 to -10dB
$R^2\theta_{az}\theta_{el}$	86dB	94dB
Duty ratio	−10dB	−10dB
Beam weighting	0dB	-5dB
	-------------	-------------
σ_c	56 to 66dB	59 to 69dB

Using these areas in place of target cross-section in the calculation of section (i), we thus conclude that mainlobe clutter will be within the range indicated by the figures below.

CNR	-23 to -13dB	-39 to -29dB.

(iii) JAMMER TO NOISE RATIO (JNR)

$$JNR = \frac{P_j G_j \lambda^2}{16\pi R^2}\left(\frac{1}{LKTFBa}\right)$$

where P_j is jammer power and G_j is the gain of its antenna. Two jammers are considered. Both emit continuous noise over the 200MHz agile bandwidth of the radar. One is a powerful ground-based system that would normally be outside the look-direction of the radar. It has a mean power of 100kW and an antenna gain of 50dB. The worst case is when it is directly below a 1000km-altitude radar. The other is an airborne stand-off or self-screening jammer which can be in the look-direction of the radar. It has a mean power of 1kW and an antenna gain of 20dB. At its closest it might be 1200km from a 1000km-altitude radar.

	Ground-based		Airborne	
	+dB	-dB	+dB	-dB
P_j	50		30	
G_j	50		20	
$\lambda^2/4$		18		18
4π		11		11
R^2		120		122
L		10		13
kTF	201		201	
Ba		83		83
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	+301	-242	+251	-247
JNR	59dB		4dB	

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Abstract

We examine the system design issues associated with the use of adaptive jammer rejection in the context of possible specifications for a space-based surveillance radar. We show that the adaptive nulling requirements of the system cannot be considered in isolation. Adaptive processing has implications for the entire system, from design of the antenna array, through to the choice of ADCs and the requirements of subsequent coherent integration and detection processing. Three different classes of adaptive processing architecture are considered: the sidelobe canceller, the fully adaptive array, and the generalised sidelobe canceller. These are shown to achieve different trade-offs between inherent complexity, requirements upon the system specification, flexibility and performance.

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